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On the Effects of Estimation Error and Jitter in Ultra-Wideband Communication

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ON THE EFFECTS OF ESTIMATION ERROR AND JITTER IN ULTRA-WIDEBAND COMMUNICATION

A Thesis
Presented to
the Graduate School of
Clemson University

In Partial Fulfillment
of the Requirements for the Degree
Master of Science
Electrical Engineering

by
Greta S. Grizzard
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Accepted by:
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ABSTRACT

The opening of the 3.6 – 10.1 GHz frequency spectrum below the “noise-floor” by the FCC in 2002 has made possible the prospect of reusing this frequency spectrum through ultra-wideband (UWB) communication. In this thesis, we compare the performance of several UWB systems in the presence of estimation error and jitter. We then develop two alternative decision schemes to combat the effect of jitter in the UWB system. Numerical results show that one of the schemes provides significantly better performance in the presence of severe jitter than maximal ratio combining and minimal degradation of performance if jitter is not present. A generalized maximal ratio combining decision scheme to combat the presence of estimation error is also proposed. It is shown that the generalized scheme outperforms traditional maximal ratio combining.
DEDICATION

In loving memory of my grandfather D.C. Scott.
ACKNOWLEDGMENTS

I thank my thesis advisor Dr. Carl Baum, for his guidance during the course of this research. I would also like to thank Dr. Wilson Pearson and Dr. John Komo for serving as members of my committee.

I would like to thank to my parents, Jane and Leslie Grizzard, for their constant support and encouragement. I am especially grateful for my grandmother, Sarah Scott, whose faithful prayers have made this accomplishment possible. Finally, and most importantly I give my greatest thanks to my Heavenly Father, for His abundance of grace, comfort, strength, mercy and blessings – the greatest of which is his Son, my Savior, Jesus Christ.

“For the LORD gives wisdom; from His mouth come knowledge and understanding” – Proverbs 2:6
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Ultra wideband (UWB) communication has recently gained much attention for its potential to reuse areas of the frequency spectrum by operating under the noise floor of existing narrowband systems. In 2002, the Federal Communications Commission (FCC) opened the 3.6 - 10.1 GHz frequency spectrum for “sub-noise-floor” UWB communication. Included in the FCC requirements is a spectral “mask” under which any given UWB device must operate. The power spectral density of a UWB signal must be extremely low, but because power is the product of the spectral density and bandwidth, the extremely large bandwidth makes communication possible. UWB communication systems are defined as using a bandwidth greater than 500MHz or having a fractional bandwidth greater than 0.2. Fractional bandwidth is defined as $B/f_c$, where $B$ is the bandwidth and $f_c$ is the center frequency $[1]$. Because of their extremely wide bandwidths, UWB systems have high multipath resolvability but are susceptible to frequency-selective fading.

Most UWB transmission schemes can be divided into three general categories: (1) orthogonal frequency division multiplexing ultra wideband (OFDM-UWB), (2) direct-sequence ultra wideband (DS-UWB), and (3) time-hopping.

$[1]$ Specifically, $B = f_H - f_L$ and $f_c = \frac{1}{2}(f_L + f_H)$ where $f_H$ and $f_L$ are the upper and lower frequencies for which the power spectral density must be at least 10 dB below the peak.
ultra wideband (TH-UWB). OFDM-UWB is characterized by the division of
the UWB frequency band into subbands which are used separately to transmit
data using multiple carrier frequencies. Both DS-UWB and TH-UWB systems
can be designed to operate using carrierless pulses. For DS-UWB, signals are
created by applying a spreading sequence, whereas for TH-UWB, the signal is
created by using a series of time shifts. A more detailed description of these
methods along with their distinct advantages and drawbacks is given below.

OFDM-UWB is a multicarrier system which divides the UWB spectrum
into roughly 500 MHz subbands [2, 3, 4, 5, 6]. This allows for the division of
the designated UWB spectrum into as many as 14 subbands. However, most
current proposals use 3 bands in the 3-5 GHz region of the available UWB
spectrum [3]. An example of OFDM-UWB given in [2] uses a frequency-
hopped pulse train of length N to transmit each BPSK information bit. Each
pulse within the train is shifted from a baseline carrier frequency by a frequency
\( c(n)/T_c \), where \( c(0), c(1), \ldots, c(N-1) \) is a pseudorandom sequence consisting of
a permutation of the integers 0, 1, ..., \( N - 1 \). The advantages of OFDM-UWB
include the ability to handle users with different data rates, the ability to
eliminate multi-user interference (MUI), and the ability to implement multi-
user detection relatively easily by taking only the FFT in the receiver [2].
OFDM-UWB also suffers several disadvantages including the need to provide
fast carrier generation (< 9.5 ns) [3], the need for users to be synchronous [2],
the need for the bandwidth to be selected to avoid narrowband interferes [5],
and the occurrence of poor performance when subject to deep fading [6].

DS-UWB is another modulation technique considered for UWB communication systems. It operates similarly to conventional wideband direct sequence spread spectrum. However, the pulse shape must be designed to give a system with at least 500 MHz bandwidth. This requires extremely short chip durations. For DS-UWB multiple access capability, each user is assigned a user-specific psuedonoise (PN) sequence. The elements of the data and spreading sequences are usually binary [7, 8, 9]. The advantages of DS-UWB include the ability to achieve data rates of up to 1 Gbps [7], low peak-to-average power ratios [9], robustness to multiple-access interference [8, 9] and the ability of users to transmit asynchronously [9]. The primary obstacle in designing an effective DS-UWB system is overcoming the effects of intersymbol interference (ISI) due to the time-dispersive nature of the UWB channel [5] and multipath fading [7].

TH-UWB (or time hopping impulse radio) was the first modulation scheme proposed for UWB communications and is still a valid method that deserves consideration. With TH-UWB each data bit is transmitted at a time that is offset by a pseudorandom sequence of delays. This time-hopping sequence can provide for multiple users and can be used to shape the frequency spectrum of the system. Whereas both OFDM and DS-UWB systems transmit pulses continuously or nearly continuously, TH-UWB systems typically have a low duty cycle. That is, if $T_f$ denotes length of time between information symbols,
called the frame duration, and $T_c$ denotes the length of the transmitted pulse, called the chip duration, then it is assumed that $T_c$ is much less than $T_f$ [10]. TH-UWB modulation has several advantages including high path resolution in the presence multipath fading, a flat noise-like spectrum, carrierless transmission, and low power requirements [10]. The principle difficulties in using TH-UWB include the effects of timing jitter, the need for precise synchronization between the transmitter and receiver, and the lower data rate compared to that of DS-UWB.

In this thesis we focus on TH-UWB. In particular we are interested in the effect of channel estimation error and timing jitter on the effectiveness of UWB communications. We seek to implement a receiver structure that gives good performance in the presence of imperfect tap weights. In the case of timing jitter, we examine the extent to which timing error degrades performance. Because the DS-UWB waveform has nearly zero autocorrelation beyond an offset of a chip, many of the conclusions of our work can be applied to DS-UWB as well. In contrast, our results have minimal application to OFDM-UWB systems.

The RAKE receiver is considered for reception of UWB signals in this work because it allows recombination of the multipath diversity using information from large numbers of paths. The RAKE receiver is widely accepted as an appropriate receiver structure for TH-UWB and DS-UWB signals. However, because of the high multipath resolution inherent to UWB, many RAKE fin-
gers are required to collect a high percentage of the transmitted signal energy.

There are three primary types of RAKE receiver described in the literature: the A-RAKE, S-RAKE, and P-RAKE. The A-RAKE (all-RAKE) uses all of the incoming multipath components [11, 12]. If $\tau_{\text{max}}$ is the largest possible multipath delay and $T_c$ is the time offset between adjacent taps, the A-RAKE must have $\tau_{\text{max}}/T_c$ taps. This number may be unacceptably large for UWB systems. The S-RAKE (selective-RAKE) chooses the $K$ “best” multipath components for combining. Details of path selection for the S-RAKE are discussed in Chapter 2. The S-RAKE is still cumbersome in that it requires the receiver to track all $\tau_{\text{max}}/T_c$ finger weights in order to determine which are best [11, 12]. However, the S-RAKE saves resources in the combining stage because only $K$ tap outputs are combined. Furthermore, depending on the technique used for combing tap outputs, the S-RAKE may outperform the A-RAKE because less noise is added in the system. Finally, the P-RAKE (partial-RAKE) combines the first $N$ RAKE tap outputs. This RAKE receiver requires the least resources in terms of tracking the multipath component arrivals [11, 12]. However, because it takes the first $N$ arrivals without regards to the signal strength in these taps, the P-RAKE generally does not perform as well as the S-RAKE with the same number of taps.

Another aspect of RAKE receiver design is the method of combining the tap outputs. Maximal Ratio Combining (MRC) is ideal in the absence of noise and estimation error. However, because of the presence of noise as well
as the inability to perfectly estimate the needed parameters of each path, other combining methods have been explored. With Equal Gain Combining (EGC), all taps that receive a multipath component are weighted with equal magnitude and combined according the estimated polarity of the received signal [13]. Other methods for combining the RAKE fingers include minimum mean square error (MMSE) [14, 15, 16] combining, Generalized Selection Combining (GSC) [17], and combining based on locally optimal detection theory [18]. MMSE-RAKE combining seeks to set the tap combining weights in such a manner as to minimize the MSE between the desired output and the received output at every tap [14]. GSC schemes choose taps which have instantaneous signal-to-noise ratio exceeding a fixed absolute (or normalized) threshold [17]. Additionally, the GSC can be combined with an adaptive threshold for the Log-Likelihood Ratio test to account for noise level, channel estimation errors, and the number of RAKE tap outputs available [17]. In [18] locally optimal detection theory is used to obtain combining rules for a RAKE receiver for UWB. The receivers are designed to mitigate an impulsive noise environment, a model that can be used to approximate the effects of multiple-access interference. In their non-Gaussian noise environment, these receivers outperform MRC combining.

In this thesis, the focus is on the use of the S-RAKE and MRC combining, because previous work shows it provides good performance without excessive complexity. For the purpose of comparison, we also consider several alternatives in Chapter 3.
In order to implement MRC detection, RAKE tap weights must be set to “match” the channel. Estimation error is the difference in the expected tap weight and the actual value needed on a tap to implement MRC detection. Estimation error is a result of the fact that the impulse response of the channel is initially unknown. Noise impedes the estimation process and results in error even if multiple pilot symbols are used to aid in estimation. Estimation error may also arise if the channel characteristics vary over time. In addition, estimation error can result from timing jitter in the system. One of the purposes of this thesis is to better understand the effects of estimation error on the performance of UWB systems.

The effects of estimation error on UWB systems has received some attention in previous work. Both [19] and [13] examine the effect of imperfect tap weights on the BER of the UWB system with BPSK modulation. A Nakagami fading channel model is used in [19], and the error rate is determined for the MRC P-RAKE and S-RAKE in the presence of noise and estimation error. In this case estimation error is modeled as a Gaussian random variable with mean zero and variance dependent on the signal to noise ratio and and inverse of the number of successive sounding symbols averaged to calculate the tap combining weights for the RAKE receiver. Although the performance improves with increasing the number of taps for both the S-RAKE and P-RAKE receivers, the opportunity cost, in terms of improved BER performance, of adding additional taps is greater in the case of imperfect estimation because
of the increased estimation error associated with the addition of each tap. The authors also conclude that basing the channel estimate on two soundings rather than a single sounding significantly improves performance with acceptably small overhead [19].

The MRC P-RAKE and the EGC P-RAKE are simulated using the IEEE 802.15.3a channel model in [13]. In the event of perfect estimation the MRC P-RAKE outperforms the EGC P-RAKE. However, in the presence of estimation error it is possible for the EGC P-RAKE to outperform the MRC P-RAKE. It is interesting to note that the authors define estimation error differently for the two receivers in [13]. For the MRC P-RAKE, estimation error is modeled as a Gaussian random variable (similar to [19]), but estimation error for the EGC P-RAKE is modeled by the probability of incorrect sign detection (i.e. deciding the multipath coefficient is negative when it is positive or deciding the multipath coefficient is positive when it is negative). It is observed that for a probability of incorrect sign detection of $10^{-3}$ there is no noticeable performance degradation in the EGC P-RAKE. Furthermore, it is claimed that the EGC P-RAKE with probability of incorrect sign detection of $10^{-1}$ outperforms the MRC P-RAKE with signal-to-noise ratio of 10 dB [13].

The estimation error which arises from estimating timing jitter can be caused by several factors, some random and some deterministic in nature. The random jitter is considered to be chiefly a result of lack of synchronization between the clock in the receiver and transmitter. In [20], a worst-case
approach is taken to analyze the effects of timing jitter. Performance is investigated in a Rayleigh fading channel in the presence of normally distributed timing jitter. Analytical results for the bit error rate are determined for binary phase-shift keying (BPSK) and pulse position modulation (PPM) with MRC RAKE reception. In [21], the effects of normally distributed timing jitter is also considered. However, in this case performance is for binary and 4-ary modulation systems and includes the effects of multiple user interference. The primary focus is the determination of how an increase in the variance of the timing jitter leads to a decrease in number of users (throughput). The system modeled does not use a RAKE receiver, but relies on the correlation of the received signal with $M$ template signals for a single symbol duration.

In this thesis, we examine the effect of estimation error on a TH-UWB system with antipodal signaling, S-RAKE reception, and MRC decision making over the suite of IEEE 802.15.3a channel models. The focus of this work is the effect of imperfect channel estimation on the the UWB system caused by noise in the system and jitter. Furthermore, new combining schemes are developed that combat the effects of estimation error. These new techniques are shown to outperform traditional MRC techniques.

The remainder of this work is organized as follows. Chapter 2 describes the system model. Chapter 3 examines the performance of a UWB system with antipodal signaling, S-RAKE receiver, and MRC decision making, and compares this performance to several other TH-UWB systems. Chapter 4
presents a S-RAKE modification that is shown to have better performance in a system prone to error due to jitter. Chapter 5 presents an additional modification of S-RAKE combining that is shown to have better performance in the presence of Gaussian estimation error. Finally, Chapter 6 offers the results and conclusions of this research.
CHAPTER 2
SYSTEM DESCRIPTION

In this chapter the transmission scheme, channel model, receiver model, and combining scheme are defined.

2.1 Transmission Scheme

A single-user system which employs antipodal signaling is presented. The transmitted signal for this system has continuous-time representation

\[ x(t) = \sqrt{E_b} \sum_{i=0}^{M-1} (-1)^{d_i} s(t - iT), \]

where \( d_i \) is the \( i \)th data bit, \( M \) is the number of transmitted data bits, and \( T \) is the delay between data bits. The system has base pulse \( s(t) \), where \( s(t) \) can be any continuous waveform that meets the following criteria: (1) the pulse has duration less than or equal to \( \tau_c \); i.e., \( s(t) = 0 \) for \( t < 0 \) and for \( t > \tau_c \); (2) the pulse is normalized to have unit energy, \( \int_0^{\tau_c} s^2(t) dt = 1 \) and (3) the pulse is designed in such a way as to fit the UWB spectral mask. It is assumed that \( T > \tau_c \). Therefore, \( E_b \) is the transmitted energy per bit for the system.

2.2 Channel Model

In this section the UWB channel model is presented. For completeness this section begins with a brief discussion of the early work done in detecting
the channel, as well as key characteristics of the UWB channel which should be reflected in the channel model in order to gain accurate insight into the behavior of a UWB system through simulation. This discussion is followed by an overview of the IEEE 802.15.3a channel model.

2.2.1 Early UWB Channel Models

Although many experiments have been performed and models developed for conventional narrowband systems, these models are inadequate for use with UWB systems because the UWB channel has a much greater multipath resolution capability than narrowband communication systems [22, 23]. There are two principle methods by which a channel may be observed, time-domain techniques and frequency-domain techniques. The time-domain technique requires transmitting a very short (< 1 ns) pulse and sampling at the receiver to recover the impulse response directly [24]. Although this method generally provides for easy measurement of variations in the channel [24], it may be difficult to create a pulse short enough to gather sufficient information to obtain a reliable model. To estimate the channel using frequency-domain sounding, a known sinusoidal signal is transmitted for each frequency step and information about the magnitude and phase of the received signal is collected. From this data, the Inverse Fourier Transform (IFT) is calculated to give the impulse response of the channel [25].

There are several characteristics of the UWB channel which must be re-
lected in the statistical model in order to perform simulations which accurately
demonstrate the performance of UWB technology. Key characteristics include
the power delay profile, arrival times, RMS delay spread, fading, path loss, and
shadowing [25, 26]. The power delay profile of the channel is a description of
the decay in amplitude of successively received multipath components. Both
the exponential distribution function and the double exponential distribution
function (i.e., two exponential distribution functions; the first for the arrival
of groups of components called clusters, and the second for the arrival of the
members of the groups called rays) have been used to describe this behavior
[25]. The inter-arrival times have been deemed to take the form of a negative
exponential distribution for the UWB channel [25]. The RMS delay spread is
the standard deviation of the delay of received multipath components weighted
proportionally to the energy in each of the received multipath components [27]
and is well-modeled as normally distributed in the observed UWB channel [25].
Fading describes the variation in received power over a local area. Fading has
been modeled using a variety of distributions including Rican, Rayleigh, and
Gamma. However, the lognormal distribution has been determined to give
the best fit in both the line-of-sight (LOS) and non-line-of-sight (NLOS) cases
[25]. Path loss is defined as the average power received as a function of dis-
tance between the transmitter and receiver. If this distance is $d$, the received
power is often modeled as proportional to $d^{-\alpha}$, where $\alpha$ is called the path-loss
exponent. Shadowing can be defined as the variation in signal power about
its mean value [26]. The lognormal distribution is also a good fit for modeling shadowing caused by path loss [25].

Several groups have previously worked to develop appropriate statistical UWB models that are simple to use for computer simulations. Some of the statistical models previously proposed include the Nakagami model [28], the $\Delta - K$ model [29], the tapped delay Rayleigh fading model [30], and the Selah-Valenzuela (S-V) model [31].

### 2.2.2 IEEE 802.15.3a UWB Channel Model Suite

The IEEE 802.15.3a suite of channel models have been designed to accurately represent the key characteristics of the UWB channel and to separately model four types of short-range UWB channels. Channel Model 1 (CM1) is a line-of-sight (LOS) channel in the 0–4 m range. Channel Model 2 (CM2) is a non-line-of-sight (NLOS) channel in the 0–4 m range. Channel Models 3 and 4 (CM3 and CM4) are also NLOS channels and represent a 4–10 m range with RMS delay spread 14.28ns and an “extreme NLOS multipath channel” with 25 ns RMS delay spread, respectively [32]. In this work the IEEE 802.15.3a channel models [33, 24, 32] are normalized to have unit energy. Thus, the discrete-time impulse response of the channel is given by

$$h(t) = \sum_{\ell=0}^{L} \sum_{k=0}^{K} \alpha_{k,\ell} \delta(t - T_\ell - \tau_{k,\ell}),$$

where the $\alpha_{k,\ell}$ are the multipath components of the channel and have a double exponential distribution (described below), $T_\ell$ is the cluster arrival time, and
\( \tau_{k,\ell} \) is the ray arrival time relative to the \( \ell \)th cluster arrival \((\tau_{0,\ell} = 0)\). The double exponential distribution of the channel implies that \( T_{\ell} \) and \( \tau_{k,\ell} \) have exponential densities

\[
P(T_{\ell} \mid T_{\ell-1}) = \Lambda \exp[-\Lambda(T_{\ell} - T_{\ell-1})], \quad T_{\ell} > T_{\ell-1}\]

and

\[
P(\tau_{k,\ell} \mid \tau_{(k-1),\ell}) = \lambda \exp[-\lambda(\tau_{k,\ell} - \tau_{(k-1),\ell})], \quad \tau_{k,\ell} > \tau_{(k-1),\ell},
\]

where \( \Lambda \) is the cluster arrival rate and \( \lambda \) is the ray arrival rate (the arrival rate within each cluster). The multipath channel coefficients are given by

\[
\alpha_{k,\ell} = \frac{p_{k,\ell} \xi_{\ell} \beta_{k,\ell}}{\sqrt{\sum_{\ell=0}^{L} \sum_{k=0}^{K} (p_{k,\ell} \xi_{\ell} \beta_{k,\ell})^2}}
\]

where each \( p_{k,\ell} \) is +1 or -1 with equal probability, \( \xi_{\ell} \) is a fading factor for the \( \ell \)th cluster, and \( \beta_{k,\ell} \) is a fading factor for the the \( k \)th path relative to the \( \ell \)th cluster. The behavior of the amplitude of the multipath components is described by

\[
20 \log_{10}(\xi_{\ell} \beta_{k,\ell}) \propto \text{Normal}(\mu_{k,\ell}, \sigma_1^2 + \sigma_2^2),
\]

where

\[
\mu_{k,\ell} = \frac{-10T_{\ell}/\Gamma - 10\tau_{k,\ell}/\gamma}{\ln(10)} \left( \frac{\sigma_1^2 + \sigma_2^2}{20} \right) \ln(10)
\]

and \( \sigma_1 \) and \( \sigma_2 \) are the standard deviations of the densities used to model the effects of the lognormal fading in clusters and rays, respectively. The values for the constants in the equations above are given in Table 2.1 for each of the four channel models (CM1 – CM4) [32].
Table 2.1: IEEE 802.13.5a UWB Channel Model Parameters

<table>
<thead>
<tr>
<th>Model Parameters</th>
<th>CM1</th>
<th>CM2</th>
<th>CM3</th>
<th>CM4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Λ [1/ns] (cluster arrival rate)</td>
<td>0.0233</td>
<td>0.4</td>
<td>0.0667</td>
<td>0.0667</td>
</tr>
<tr>
<td>λ [1/ns] (ray arrival rate)</td>
<td>2.5</td>
<td>0.5</td>
<td>2.1</td>
<td>2.1</td>
</tr>
<tr>
<td>Γ (cluster decay factor)</td>
<td>7.1</td>
<td>5.5</td>
<td>14.00</td>
<td>24.00</td>
</tr>
<tr>
<td>γ (ray decay factor)</td>
<td>4.3</td>
<td>6.7</td>
<td>7.9</td>
<td>12</td>
</tr>
<tr>
<td>σ₁ [dB] (stand. dev. of cluster</td>
<td>3.3941</td>
<td>3.3941</td>
<td>3.3941</td>
<td>3.3941</td>
</tr>
</tbody>
</table>
lognormal fading term in dB)          |
| σ₂ [dB] (stand. dev. of ray          | 3.3941  | 3.3941 | 3.3941 | 3.3941 |
lognormal fading term in dB)          |
| σₙ [dB] (stand. dev. of              | 3       | 3      | 3      | 3      |
lognormal fading term for total       |
multipath realizations in dB)         |

As presented in [10, 16, 17], the IEEE802.15.3a channel model may be represented as

\[ h(t) = \sum_{j=0}^{N} h_j \delta(t - j\tau_c), \]

(2.4)

where \( h_j \) is the sum of all \( \alpha_{k,\ell} \) terms for which \((j - 1)\tau_c < (T_\ell + \tau_{k,\ell}) \leq j\tau_c\).

Note that some of the \( h_j \)s are likely to be zero; i.e., there are no \( \alpha_{k,\ell} \) terms in this time interval. In addition, \( N \) is chosen such that \( N\tau_c \) is long enough to capture the entire channel with high probability.
2.3 Receiver Description

A RAKE receiver is used to collect the multipath components of the channel. The weights of the RAKE tap outputs are a function of an estimate of the channel and the combining scheme. The channel can be estimated as having impulse response

\[ \hat{h}(t) = \sum_{j=0}^{N} \hat{h}_j \delta(t - j\tau_c) \]  

(2.5)

where \( \hat{h}_j \) is the estimate of the channel at the \( j \)th RAKE tap. In this section, the methods for obtaining the channel estimates and the tap weights are developed.

The received signal for a single transmitted bit \( d_i \) in the presence of noise \( n(t) \) is given by

\[
r(t) = [x * h](t) + n(t)
= \sqrt{E_b} \sum_{i=0}^{M-1} \sum_{j=0}^{N} h_j(-1)^{d_i}s(t - iT - j\tau_c) + n(t),
\]

where the \( h_j \)s are as defined in Eq. 2.4, \( T \) is the delay between data bits, and \( \tau_c \) is the pulse duration. We assume that \( n(t) \) is additive white Gaussian noise (AWGN) with spectral density \( N_0/2 \).

Suppose a known data sequence of length \( N_{cs} \) is used to estimate the channel, i.e., \( d_0 = d_1 = \ldots = d_{N_{cs}-1} = 1 \). Consider

\[
\hat{h}_{j,i} = \frac{1}{\sqrt{E_b}} \int_{iT+j\tau_c}^{iT+(j+1)\tau_c} r(t) s(t - iT - j\tau_c) \, dt,
\]

\[ 0 \leq i \leq N_{cs} - 1; \]
i.e., $\hat{h}_{j,i}$ is matched to the base pulse $s(t)$. Assuming $T > N\tau_c$,

$$
\hat{h}_{j,i} = \frac{1}{\sqrt{E_b}}[\sqrt{E_b}h_j + \int_{iT+j\tau_c}^{iT+(j+1)\tau_c} n(t)s(t - iT - j\tau_c) \, dt]
$$

$$
= h_j + N_{j,i},
$$

where $N_{j,i}$ is normally distributed with mean zero and variance $N_0/2$. We obtain a single estimate for $h_j$ with the average

$$
\hat{h}_j = \frac{1}{Ncs} \sum_{i=0}^{Ncs-1} \hat{h}_{j,i}.
$$

(2.6)

It is straightforward to show that $\hat{h}_j$ is normally distributed with mean $h_j$ and variance $N_0/(2N_{cs})$.

We now describe how the tap weights in the S-RAKE receiver are obtained. Only the $L_c$ taps with largest estimated magnitude are used for combining. Therefore, the S-RAKE taps, $\hat{h}_{j,s}^*$, $0 \leq j \leq N$, are defined as

$$
\hat{h}_{j,s}^* = \begin{cases} 
\hat{h}_j & \text{if } j \in S \\
0 & \text{else},
\end{cases}
$$

(2.7)

where $S$ is the set of indices corresponding to the $L_c$ taps with largest magnitude.

For the S-RAKE receiver with MRC decision making, the detected bit $\hat{d}_i$ is given by

$$
\hat{d}_i = \begin{cases} 
0 & \text{if } \sum_{j \in S} \hat{h}_{j,s}^* y_{j,i} \geq 0 \\
1 & \text{if } \sum_{j \in S} \hat{h}_{j,s}^* y_{j,i} < 0
\end{cases}
$$

(2.8)
where $y_{j,i}$ is the output matched to the base pulse $s(t)$ corresponding to the $j$th time interval of length $\tau_c$ for the $i$th data bit; i.e.,

$$y_{j,i} = \frac{1}{\sqrt{E_b}} \int_{iT+j\tau_c}^{(iT+(j+1))\tau_c} r(t) * s(t-iT-j\tau_c) \, dt. \quad (2.9)$$
CHAPTER 3
EFFECTS OF ESTIMATION ERROR

In this chapter the performance of a UWB system with antipodal signaling, S-RAKE reception, and MRC decision making is examined. The performance of the system in the presence of estimation error is presented, and is compared to that of other choices for signaling and reception of UWB systems. Included are systems employing various combinations of pulse position modulation (PPM), P-RAKE reception, and equal-gain combining (EGC), and individual features of the initial system. Finally, the effects of receiver parameter value on performance is investigated.

3.1 Performance of Antipodal Signaling with MRC S-RAKE

In this section the performance of the of the UWB system with BPSK modulation, S-RAKE ($L_c = 20$), and MRC decision making is considered. Specifically, the performance of the system is analyzed from two perspectives: versus $E_b/N_0$ for a fixed number of channel soundings, $N_{cs}$ (1, 2, 10) and versus the number of channel soundings for a fixed $E_b/N_0$ (10 dB, 15 dB). Fig. 3.1 presents the performance range achievable for 1, 2, and 10 channel soundings versus $E_b/N_0$. From this figure it is clear that sounding the channel 10 times, instead of only once, improves the performance of the system between approximately 5 and 6 dB, depending on the channel parameters. It is also interesting
to note the increase in performance gained by sounding the channel twice. The system experiences an increased performance gain of roughly 2 dB, depending on the channel.

Fig. 3.2 shows the performance of the S-RAKE receiver with $L_c = 20$ taps as a function of the number of channel soundings, $N_{cs}$. From this figure it is clear that even a small number of channel soundings brings a large gain in performance of the system, especially at high signal-to-noise ratios. However, the system also encounters the law of diminishing returns for large numbers of channel soundings; that is, the system experiences little improvement for additional channel soundings above a given threshold. Fig. 3.2(a) shows that, for $E_b/N_0 = 10$ dB, beyond $N_{cs} = 30$ CM1 shows almost no improvement with additional channel soundings. Likewise, CM4 shows little improvement
Figure 3.2: Performance of MRC Receiver with Channel Estimation Error versus Number of Channel Soundings, Antipodal Signaling, S-RAKE Reception ($L_c = 20$)
beyond $N_{cs} = 50$, at $E_b/N_0 = 10$ dB. At the higher SNR of $E_b/N_0 = 15$ dB, the increased performance caused by only a few channel soundings is remarkable. Sounding the channel 10 times (Fig. 3.2(b)) shows that a probability of bit error ($P_b$) of less than $10^{-5}$ is achievable for CM1 – CM3 and less than $10^{-4}$ for CM4.

The increase in performance from sounding the channel repeatedly comes at the cost of additional delay during the transmitter and receiver acquisition and reacquisition periods. The value of this additional delay is $(N_{cs} - 1)T$.

3.2 Pulse Position Modulation and Demodulation

We consider pulse position modulation (PPM) as an alternative to antipodal signaling. For PPM, the transmitted signal is given by

$$x_{PPM}(t) = \sqrt{E_b} \sum_{i=0}^{M-1} s(t - d_i \tau_c - i(T + \tau_c))$$

where $d_i$ is the $i$th data bit, $\tau_c$ is the PPM timing offset and is equal to the chip time, and $T$ is the same delay between bit transmissions as that used for antipodal signaling.

Then for a channel described as

$$\hat{h}(t) = \sum_{j=0}^{N} h_j \delta(t - j\tau_c)$$
as in Eq. 2.4, the received PPM signal is given by

\[
r_{PPM}(t) = [x_{PPM} * h](t) + n(t)
\]

\[
= \sqrt{E_b} \sum_{i=0}^{M-1} \sum_{j=0}^{N} h_j s(t - j\tau_c - d_i\tau_c - iT) + n(t)
\]

\[
= \sqrt{E_b} \sum_{i=0}^{M-1} \sum_{j=0}^{N} h_j s(t - (j + d_i)\tau_c - iT) + n(t).
\]

For PPM signaling the detected bit \( \hat{d}_i \) is given by

\[
\hat{d}_i = \begin{cases}
0 & \text{if } \sum_{j \in S} \hat{h}_j y_{j,i} \geq 0 \\
1 & \text{if } \sum_{j \in S} \hat{h}_j y_{j,i} < 0
\end{cases}
\] (3.1)

where \( y_{j,i} \) is the output matched to the difference of the two possible signals; i.e.,

\[
y_{j,i} = \frac{1}{\sqrt{E_b}} \int_{iT + j\tau_c}^{iT + (j+1)\tau_c} r_{PPM}(t)[s(t - iT - j\tau_c) - s(t - iT - (j + 1)\tau_c)] \, dt.
\] (3.2)

### 3.3 P-RAKE Receiver

The P-RAKE receiver operates similarly to the S-RAKE receiver, except where the S-RAKE receiver sums over the \( L_c \) taps with largest magnitude, the P-RAKE receiver sums over the first \( L_c \) taps. Therefore, the taps for the P-RAKE receiver are given by

\[
\hat{h}_j^p = \begin{cases}
\hat{h}_j & \text{if } 0 \leq j < L_c \\
0 & \text{else}
\end{cases}
\]
That is, the selection set $S$ equals $\{0, 1, \ldots, L_c - 1\}$. Detection is achieved by replacing $\hat{h}_j^s$ with $\hat{h}_j^p$ in Eq. 2.8 if using antipodal signaling and Eq. 3.1 if using PPM.

The P-RAKE has a distinct advantage over the S-RAKE in the simplicity of design because the P-RAKE is only required to track the first $L_c$ tap outputs of the RAKE receiver, whereas the S-RAKE must track RAKE outputs for the entire length of the channel during the acquisition period in order to ensure it chooses the $L_c$ taps with largest magnitude. However, because the taps chosen with the P-RAKE are typically not the largest values, performance is expected to be inferior to that of the S-RAKE if both have the same number of tap outputs.

### 3.4 Equal Gain Combining

Equal Gain Combining (EGC) is a lower-complexity alternative to MRC detection. Here, all non-zero taps in the RAKE receiver are set to plus or minus one, depending on the sign of the received signal. Let $y_{j,i}$ be the signal received by the $j$th RAKE finger, given by Eq. 2.9 if antipodal signaling is used and by Eq. 3.2 if PPM is used. Then the EGC decision is given by

$$
\hat{d}_i = \begin{cases} 
0 & \text{if } \sum_{j \in S} \text{sgn}(\hat{h}_j)y_{j,i} \geq 0 \\
1 & \text{if } \sum_{j \in S} \text{sgn}(\hat{h}_j)y_{j,i} < 0
\end{cases}
$$

where $\hat{h}_j$ is equal to $\hat{h}_j^s$ for the S-RAKE receiver and $\hat{h}_j^p$ for the P-RAKE, and where $S$ is defined appropriately for the two types of receivers.
3.5 Performance Comparison of UWB Systems

From Figs. 3.3 – 3.14 it is clear that for the same receiver and decision scheme, the antipodal signaling system normally gives better performance than PPM. The amount of the advantage however, does vary across the systems. For systems operating at 10 dB with only one or two channel soundings (Figs. 3.3, 3.4, 3.6, 3.7, 3.9, 3.10, 3.12 and 3.13 (a)) the PPM systems give nearly the same performance as the antipodal signaling systems when a large number of S-RAKE fingers are used ($L_c > 60$). However, the performance of the PPM systems in these events is worse than the antipodal signaling performance attainable with fewer S-RAKE taps. Additionally, Figs. 3.5, 3.8, 3.11, and 3.14 (a) show that for $E_b/N_0 = 10$ dB with ten channel soundings antipodal signaling outperforms PPM for $L_c < 35$. In systems with $E_b/N_0 = 15$ dB, Figs. 3.3 - 3.14 (b), antipodal signaling usually outperforms PPM by more than an order of magnitude for a relatively short RAKE receiver ($L_c = 10$ to 30 in the case of the S-RAKE). Likewise, using the P-RAKE receiver the antipodal signaling systems give better performance than the PPM systems with P-RAKE receiver by an order of magnitude or more. For this reason we conclude that antipodal signaling is preferable to PPM if moderately accurate estimates of the channel can be obtained.

We now compare the performance of antipodal signaling systems with MRC and EGC decisions. If the EGC scheme will ever have and advantage over
Figure 3.3: CM1: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 1$
Figure 3.4: CM1: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 2$
Figure 3.5: CM1: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 10$
Figure 3.6: CM2: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 1$
Figure 3.7: CM2: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 2$

(a) $E_b/N_0 = 10$ dB

(b) $E_b/N_0 = 15$ dB
Figure 3.8: CM2: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 10$
Figure 3.9: CM3: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 1$
Figure 3.10: CM3: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 2$

(a) $E_b/N_0 = 10 \text{ dB}$

(b) $E_b/N_0 = 15 \text{ dB}$
Figure 3.11: CM3: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 10$
Figure 3.12: CM4: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 1$
Figure 3.13: CM4: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 2$
Figure 3.14: CM4: Comparison of Standard System Performance with Channel Estimation Error, $N_{cs} = 10$
MRC, it will be when the information received is the most unreliable, that is, at $E_b/N_0$ (i.e. 10 dB) and for systems which use few channel soundings ($N_{cs} = 1, 2$). Therefore, we again focus on Figs. 3.3, 3.4, 3.6, 3.7, 3.9, 3.10, 3.12 and 3.13 (a). From these eight figures it is clear that while EGC gives comparable performance under these circumstance, it does not outperform MRC even when the information received is unreliable. However, the discrepancy between the performance of MRC systems and EGC systems is apparent, as expected, for systems in which the received information is more reliable. For example, in Fig. 3.11(a), the MRC, antipodal signaling, S-RAKE reception system outperforms the EGC, antipodal signaling, S-RAKE reception system by roughly a factor of 2 for a S-RAKE with 20 or more taps. The performance gain for MRC over EGC is even greater for CM1 where with 10 channel soundings (Fig. 3.5(a)), the MRC, antipodal signaling, S-RAKE reception system outperforms the EGC counterpart by nearly an order of magnitude for the S-RAKE receiver with 20 or more taps. MRC, antipodal signaling, P-RAKE reception systems also outperform the EGC, antipodal signaling, P-RAKE reception systems. For these systems the improvement generally increases with increasing RAKE length. As with the S-RAKE reception systems, the improvement ranges from nearly identical performance to approximately an order of magnitude. Also note that all PPM systems with MRC also outperform their EGC counterparts.

The comparison of S-RAKE and P-RAKE receivers is the most interesting
of the three components of the UWB system. Neither system gives the best performance in all cases. In the majority of cases in the figures shown the P-RAKE receiver’s best performance is better than the S-RAKE receiver’s best performance. For the LOS channel, CM1, in Figs. 3.3 – 3.5 it is observed that for low $E_b/N_0$ and few channel soundings the P-RAKE receiver outperforms the S-RAKE receiver across all lengths. This can be attributed to the increased reliability of the early arrivals in the LOS case which may not have experienced as much reflection as in the NLOS case. In many cases the P-RAKE receiver is also capable of outperforming the S-RAKE for NLOS channels, CM2 – CM4, as well. However, this improvement in performance comes at the cost of a much larger RAKE receiver. The P-RAKE receiver requires between 40 and 70 taps in most cases to outperform the best performing S-RAKE receiver for CM2 – CM4 (Figs. 3.6 – 3.14). The S-RAKE receiver usually achieves its best performance for systems with antipodal signaling, MRC combining, and between 10 and 30 taps. Figs. 3.12-3.14 show that CM4 is the exception to this rule in that it does not attain a local minimum for $L_c < 100$. In this case, while the performance of the antipodal signaling, S-RAKE reception, MRC system improves with increasing RAKE length, the slope of this improvement is nearly flat, so a shorter RAKE sacrifices little in performance. Therefore, because of the cost and complexity of using a larger P-RAKE receiver, the smaller S-RAKE receiver is generally preferable.
3.6 Selection of System Parameters

From the arguments presented in Section 3.5, a preferred system employs antipodal signaling, the S-RAKE receiver, and MRC decision making. The parameter remaining to be considered is the length of the S-RAKE receiver, $L_c$. It is desirable to keep $L_c$ small ($< 30$) in order to reduce the complexity of the system. From Fig. 3.15 it is clearly impossible to obtain the best performance across all cases of channel model, $E_b/N_0$, and $N_{cs}$ for a single value of $L_c$, because the cases do not achieve their minimum at the same $L_c$. Moreover, because it is impossible to know the environment in which a system will be operating before implementation, $L_c$ must be chosen to give the best “overall” performance. The great emphasis is given to the systems with more channel soundings ($N_{cs} = 10$) because these systems give much better performance. Additionally, greater consideration is given to performance for $E_b/N_0 = 10$ dB than for $E_b/N_0 = 15$ dB because the antipodal signaling, S-RAKE reception, MRC systems with ten channel soundings all perform well if $E_b/N_0 = 15$ dB. Therefore, the emphasis is placed on the solid lines in Fig. 3.15(a). Looking at these lines consider S-RAKE receiver with lengths of 15, 20, and 30, as marked by the vertical lines on the plot. Comparing the performance for $L_c = 20$ and $L_c = 30$ it is observed that there is almost no performance improvement for using 30 taps instead of 20 for CM2 – CM4 and CM1 experiences a slight decrease in performance. Next, consider the performance for $L_c = 15$ and
Figure 3.15: Performance of Antipodal Siganling with MRC and Channel Estimation Error versus Length of S-RAKE Receiver
$L_c = 20$. In this case CM2 – CM4 experience a small gain in performance at $L_c = 20$ while CM1 has approximately the same performance for both receiver lengths. However, considering the antipodal signaling, S-RAKE reception, MRC systems with $N_{cs}$ when $E_b/N_0 = 15$ dB is observed that CM4 gains about a factor of 1.5 in performance for $L_c = 20$ compared to $L_c = 15$, where as CM1 – CM3 all have probability of bit error less than $10^{-5}$ for both lengths. Finally, for $N_{cs} = 1, 2$ most of the channel models experience some improvement in performance, usually a few tenths of a decibel at $L_c = 20$ compared to $L_c = 15$. Therefore, we chose the modest increase in complexity and recommend $L_c = 20$ for the parameters of the system considered here.
CHAPTER 4
EFFECTS OF JITTER ON UWB

In this chapter the effect of jitter in the UWB system and/or channel is examined. The chapter begins with a description of the modifications made to the channel model to account for jitter in the system. We then propose two techniques for mitigating jitter based on a modification of the tap weights used in MRC detection. Finally, we compare the performance of the jitter mitigation receivers with that of a traditional MRC receiver.

4.1 Modified Channel Model for Jitter

In Chapter 2 the UWB channel was defined as

\[ h(t) = \sum_{j=0}^{N} h_j \delta(t - j\tau_c). \]

Here, \( h_j \) incorporates all the multipath components that arise over a channel bin of length \( \tau_c \). Jitter results in some of these components being advanced or delayed relative to the previous or subsequent channel bin. To develop analysis to allow for the presence of jitter in the system at the time of acquisition, the “pre-jitter” impulse response of the channel, \( \tilde{h}(t) \), is defined by

\[ \tilde{h}(t) = \sum_{j=0}^{N} \tilde{h}_j \delta(t - j\tau_c). \]

We model the relationship between \( h_j \) and \( \tilde{h}_j \) by the following equation:

\[ h_j = (1 - |\rho|)\tilde{h}_j + |\rho| \tilde{h}_{j+\text{sgn}(\rho)}, \quad 0 \leq j \leq N, \]
where \(-1 \leq \rho \leq 1\), \(|\rho|\) indicates the proportion of the signal received in the adjoining bin, and \(\text{sgn}(\rho)\) indicates the whether the signal was delayed or advanced. Additionally, \(\tilde{h}_{-1} = \tilde{h}_{N+1} = 0\); that is, the channel is assumed to be clear before the arrival of the first multipath component and after the arrival of the last multipath component. Thus the channel model for \(h(t)\) with jitter can be rewritten as

\[
h(t) = \sum_{j=0}^{N} ((1 - |\rho|)\tilde{h}_j + |\rho|\tilde{h}_{j+\text{sgn}(\rho)})\delta(t - j\tau_c).\]

It is important to recognize several assumptions that have been made about the jitter present in the system in these equations. First, it is assumed that for any channel realization, the proportion of the signal which is advanced (or delayed) is constant. Second, the jitter in the system causes either a delay or advance of the received signal for a given channel realization; it is not possible for some bins to experience delays while others experience advances. Finally, it is assumed that the presence of jitter is restricted to adjacent bins.

4.2 Jitter Mitigation Receivers

One method which may be used to try and mitigate the effect jitter has on the system is to add a weighted amount of the signal received in adjacent bins to the tap weight for a given bin. This spreads the information received in each bin to its neighbors in an attempt to more accurately reflect the uncertainty in the bin location of multipath components. In this section, two methods for decision making using receiver tap weights with this additional information are
considered. The Jitter Mitigation Receiver (JMR) operates similarly to the
MRC receiver, but uses the tap weights described above. The Comparative
Jitter Mitigation Receiver (CJMR) calculates both the MRC decision statistic
and the JMR decision statistic and makes the decision based on the decision
statistic with the largest magnitude.

Specifically, for the JMR, the weight of the tap weight calculated by the
jth finger of the RAKE receiver is given by

\[
\hat{h}_j(\alpha) = \alpha \hat{h}_{j-1} + \hat{h}_j + \alpha \hat{h}_{j+1}, \quad 0 \leq j \leq N,
\]

where, \(0 \leq \alpha \leq 1\) and \(\hat{h}_{-1} = \hat{h}_{N+1} = 0\). Note that if \(\alpha = 0\), jitter is ignored
and we have \(\hat{h}_j(0) = \hat{h}_j\). For an S-RAKE with \(L_c\) taps, let

\[
\hat{h}_j^S(\alpha) = \begin{cases} 
\hat{h}_j(\alpha) & \text{if } j \in S, \\
0 & \text{else}
\end{cases}
\]

where \(S\) is the set of indices corresponding to the \(L_c\) tap weights with largest
magnitude. Then, using antipodal signaling and MRC with the new \(\hat{h}_j^S(\alpha)\) tap
weights, the receiver decides

\[
\hat{d}_i = \begin{cases} 
0 & \text{if } \sum_{j=0}^{N} \hat{h}_j^S(\alpha) y_{j,i} \geq 0, \\
1 & \text{if } \sum_{j=0}^{N} \hat{h}_j^S(\alpha) y_{j,i} < 0
\end{cases}
\]

where

\[
y_{j,i} = \frac{1}{\sqrt{E_b}} \int_{iT+j\tau_c}^{iT+(j+1)\tau_c} r(t) * s(t - iT - j\tau_c) \, dt,
\]
The Comparative Jitter Mitigation Receiver (CJMR) uses the information used to make the MRC and the JMR decisions and compares their magnitudes to get the information to make its decision. That is, let

\[ b_{i}^{JMR} = \sum_{j=0}^{N} \hat{h}_{j}^{s}(\alpha) y_{j,i} \]

and

\[ b_{i}^{MRC} = \sum_{j=0}^{N} \hat{h}_{j}^{s}(0) y_{j,i} \]

where \( \hat{h}_{j}^{s}(\alpha) \) is the tap weight of the \( j \)th RAKE tap output for the length \( L_{c} \) JMR S-RAKE and \( \hat{h}_{j}^{s}(0) \) is the tap weight of the \( j \)th RAKE tap output for the length \( L_{c} \) S-RAKE for the MRC receiver.

Then, the CJMR decides as follows

\[ b_{i}^{CJMR} = \begin{cases} 
    b_{i}^{JMR} & \text{if } |b_{i}^{JMR}| \geq |b_{i}^{MRC}|, \\
    b_{i}^{MRC} & \text{if } |b_{i}^{JMR}| < |b_{i}^{MRC}| 
\end{cases} \]

\[ \hat{d}_{i} = \begin{cases} 
    0 & \text{if } b_{i}^{CJMR} \geq 0, \\
    1 & \text{if } b_{i}^{CJMR} < 0. 
\end{cases} \]

4.3 Performance of JMR and CJMR

A system using antipodal signaling and S-RAKE reception with \( L_{c} = 20 \) taps is considered. The primary objective of this section is to evaluate the performance of both the JMR and CJMR receivers in the presence of jitter with the hope of obtaining a receiver which performs better than MRC in the presence of substantial jitter, while still making reliable decisions when jitter
Figure 4.1: CM1: Performance of JMR with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 4.2: CM2: Performance of JMR with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 4.3: CM3: Performance of JMR with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 4.4: CM4: Performance of JMR with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
is not present or is insignificant. To isolate the effects of jitter we assume that
$N_{cs}$ is sufficiently large so that the effects of non-jitter channel estimation error
is negligible. Figs. 4.1 – 4.4 (a) show the performance of the Jitter Mitigation
Receiver (JMR) for a system with $E_b/N_0 = 10$ dB. For all channel models, the
case where $\alpha = 0$ is the MRC receiver.

Figs. 4.1 – 4.4 show that if $|\rho|$ is small, the MRC outperforms the JMR,
and with a large value of $|\rho|$, the JMR outperforms the MRC. The use of
large value of $\alpha$ results in a significant hit in performance if $|\rho|$ is small, but
also provides significant gains if $|\rho|$ is large. The gains for large $|\rho|$ are even
more impressive in Figs. 4.1 – 4.4 (b) where $E_b/N_0 = 15$ dB than they are in
Figs. 4.1 – 4.4 (a) where $E_b/N_0 = 10$ dB. The non-monotonic behavior of the
$\alpha = 1.0$ receiver with $E_b/N_0 = 15$ dB in CM1 and CM2 in Figs 4.1 and 4.2 (b)
may be due to the fact that the channels are “dense” with paths and using
adjacent taps is more likely to cause destructive interference.

The Comparative Jitter Mitigation Receiver (CJMR) seeks to preserve the
gain in performance experienced by the Jitter Mitigation Receiver (JMR) in
the presence of severe jitter ($|\rho| \approx 1$) while reducing the hit taken by the JMR
when no jitter is present in the system ($\rho = 0$). Figs. 4.5 - 4.8 show the greatly
improved performance of the CJMR over the JMR in the absence of jitter.

For example, with $\rho = 0$, channel model 1, and $E_b/N_0 = 10$ dB (c.f.
Fig. 4.5(a)), the probability of bit error when for MRC is $2.1 \times 10^{-5}$ for $\alpha = 0.6$
JMR is $6.2 \times 10^{-4}$ and for $\alpha = 0.6$ CJMR is $4.7 \times 10^{-5}$, the latter is nearly
as good as MRC. Yet at $\rho = 1$ the $\alpha = 0.6$ receiver outperforms MRC by a wide margin (error probabilities of 0.165 versus 0.386). Thus, we see that the CJMR receiver does an excellent job of mitigating jitter without degrading performance significantly when jitter is absent.
Figure 4.5: CM1: Performance of CJMR and JMR Receivers with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 4.6: CM2: Performance of CJMR and JMR Receivers with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 4.7: CM3: Performance of CJMR and JMR Receivers with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 4.8: CM4: Performance of CJMR and JMR Receivers with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
CHAPTER 5
GENERALIZED MAXIMAL RATIO COMBINING FOR UWB

5.1 Motivation

In this chapter a RAKE combining scheme is proposed as an alternative to MRC detection for UWB with antipodal signaling and S-RAKE reception in the presence of channel estimation error. The MRC approach is optimal only if the channel weights are known perfectly. Because of this fact, we propose a generalization of the MRC scheme and investigate its performance in the subsequent sections.

5.2 Description of Generalized MRC Scheme

Let, $\hat{h}_j$, $0 \leq j \leq N$ be a set of RAKE output tap weights generated by $N_{cs}$ channel soundings of a given channel. Then for an S-RAKE receiver with $L_c$ nonzero tap weights, set

$$\hat{h}_j^s = \begin{cases} 
\hat{h}_j & \text{if } j \in S, \\
0 & \text{else.}
\end{cases}$$

where $S$ is the set of indices corresponding to the $L_c$ taps with largest magnitude. Then, the generalized MRC decision rule of degree $p$ for the BPSK
system is defined as

\[
\hat{d}_i = \begin{cases} 
0 & \text{if } \sum_{j=0}^{N} \left( |N_{cs} \hat{h}_j^s - y_{j,i}| \right)^p \geq \sum_{j=0}^{N} \left( |N_{cs} \hat{h}_j^s + y_{j,i}| \right)^p \\
1 & \text{if } \sum_{j=0}^{N} \left( |N_{cs} \hat{h}_j^s - y_{j,i}| \right)^p < \sum_{j=0}^{N} \left( |N_{cs} \hat{h}_j^s + y_{j,i}| \right)^p 
\end{cases}
\]

where

\[
y_{j,i} = \frac{1}{\sqrt{E_b}} \int_{iT+T_c}^{iT+(j+1)T_c} r(t) \ast s(t - iT - jT_c) \, dt,
\]

is the received signal for the \( j \)th tap output of the RAKE receiver. It can readily be shown that \( p = 2 \) is equivalent to the MRC receiver. Note that \( p \) is not restricted to be an integer but must be real and positive.

5.3 Performance of Generalized MRC Detection

Figs. 5.1 – 5.4 show the performance of the generalized detection rule for \( p = 1, 2, 3, \) and 6 as well as MRC performance (\( p = 2 \)). From these figures it is clear that the generalized MRC detection rule with \( p = 3 \) consistently outperforms MRC across all systems and \( E_b/N_0 \) considered. In contrast, \( p = 1.5 \) gives performance consistently worse than MRC across all systems and \( E_b/N_0 \) considered. For \( p = 6 \) the receiver performs well if there are few channel soundings or \( E_b/N_0 \) is small. However, as the information becomes more reliable, the \( p = 6 \) receiver degrades in performance compared to the MRC and \( p = 3 \) receivers.

The results show that the \( p = 3 \) receiver provides the greatest improvements over MRC decision making with CM1 and CM2. One reason for this
may be that a greater amount of energy is captured with largest taps with
these channel models, and thus, there is less “averaging out” of error, and
errors in estimation have a more detrimental effect on performance.
Figure 5.1: CM1: Performance of Generalized MRC with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)

Figure 5.2: CM2: Performance of Generalized MRC with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
Figure 5.3: CM3: Performance of Generalized MRC with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)

Figure 5.4: CM4: Performance of Generalized MRC with Antipodal Signaling, S-RAKE Receiver ($L_c = 20$)
CHAPTER 6
CONCLUSIONS

In this thesis we have investigated the effect of estimation error from noise and jitter on the performance of a UWB communications system. Our primary system uses antipodal signaling, S-RAKE reception, and MRC detection.

We have compared the performance of antipodal signaling versus PPM, of MRC versus EGC detection, and of S-RAKE versus P-RAKE reception. It has been shown that no single combination of transmission scheme, decision scheme, and RAKE receiver structure gave the best performance in all channel models and signal-to-noise ratios. However, we conclude the “overall best” system uses antipodal signaling, S-RAKE reception, and MRC decision making. Additionally, we choose to use a length $L_c = 20$ S-RAKE receiver, because this length gives good performance without excessive complexity.

We have also presented two alternative methods of combining to combat jitter in the UWB communications system. The Jitter Mitigation Receiver (JMR) adds a fraction of the signal received by both the preceding and the succeeding tap outputs to each RAKE tap. This combining scheme gives improved performance in the presence of jitter, especially when the jitter is severe, but it does not perform well if jitter is absent. We have also considered the Comparative Jitter Mitigation Receiver (CJMR) in which we compare the decision statistic for the MRC and the JMR and make the decision based on
the statistic with the largest magnitude. This decision scheme gives increased performance in the presence of jitter and sacrifices little in performance when jitter is absent.

We have also developed and presented a generalization of the MRC decision rule as an alternative means of combining to combat estimation error due to noise. We have shown that the generalized MRC receiver with $p = 3$ gives better performance than standard MRC ($p = 2$).

Opportunities for future work involve investigating the performance of combining the polynomial receiver and the CJMR for the situation in which both estimation error due to noise and jitter are present. Additionally, it may be beneficial to consider a scheme in which the number of channel soundings is dependent on the mean and variance of the received signals at the RAKE tap outputs from the first few channel soundings.
REFERENCES


